Interference Suppression in WCDMA Downlink by Symbol-Level Channel Equalization

Kari Hooli, Markku Juntti
University of Oulu, Centre for Wireless Communications
P.O. Box 4500 FIN-90014 University of Oulu, Finland
Tel. +358 8 553 2883, Fax +358 8 553 2845
email: kari.hooli@ee.oulu.fi

ABSTRACT

The main air-interface in the 3rd generation cellular communications is wideband CDMA (WCDMA). For WCDMA downlink, multiple access interference (MAI) suppression by symbol-level equalization has proven to be problematic due to long spreading sequences employed in downlink. To circumvent the problem, receivers based on channel equalization at chip level have been proposed to ensure adequate performance even with a high number of active users. In this paper, a symbol-level equalizer is introduced. The equalizer provides significant complexity reduction over the chip-level equalizer without compromising the performance. The numerical results show significant performance improvement over the conventional Rake receiver with a high number of active users in a Rayleigh fading multipath channel.

I. INTRODUCTION

The air interface of universal terrestrial radio access (UTRA), the most important 3rd generation cellular mobile communications standard, is based on wideband code-division multiple-access (WCDMA). The downlink capacity is expected to be more crucial than the capacity of the uplink due to the asymmetric capacity requirements, i.e., the downlink direction should offer higher capacity than the uplink [1]. Therefore, the employment of efficient downlink receivers is important. The performance of the conventional Rake receiver has been reported to be insufficient with a high number of active users [2]. In order to avoid performance degradation near-far resistant (or multiuser) receivers can be used. However, optimal receivers in direct-sequence CDMA (DS-CDMA) systems are far too complex for practical implementations. Therefore several suboptimal receivers feasible for practical implementations have also been proposed. E.g., in [3]-[4] linear minimum mean squared error (LMMSE) receivers were presented for DS-CDMA systems. The adaptive versions of the symbol level LMMSE receivers rely on cyclostationarity of multiple access interference (MAI), and thus require periodic spreading sequences with very short period. Hence they can not be directly applied on the WCDMA downlink.

In a synchronously transmitted downlink employing orthogonal spreading codes MAI is mainly caused by multipath propagation. Due to the non-zero cross-correlations between the spreading sequences with arbitrary time shifts, there is interference between propagation paths (or Rake fingers) after the despreading causing MAI. If the received chip waveform, distorted by the multipath channel, is equalized prior to the correlation by the spreading code or matched filtering, there is only a single path in the despreading. With orthogonal spreading sequences the equalization effectively retains, to some extent, the orthogonality of users lost due to the multipath propagation, thus suppressing MAI. Since the channel equalization does not depend on the period of spreading sequences, it can also be applied in systems using long spreading sequences. Such a receiver, discussed e.g. in [5]-[7], consists of a linear equalizer followed by a single correlator and a decision device. Several adaptive versions of chip-level channel equalizers have been presented e.g. in [8]-[10] and in the references therein.

In this paper, a novel receiver performing the channel equalization at symbol-level is presented. Through the symbol-level signal processing, the receiver achieves significant complexity reduction over the chip-level equalizer without compromising the performance. The receiver is based on the approximations done for the LMMSE receiver on the chip-level, and the connection between the presented receiver and LMMSE receiver is the same as with the chip-level equalizers.

The rest of the paper is organized as follows. The system model is defined in Section II, and the receiver is defined in Section III. The performance of the equalizer is evaluated and compared to the conventional Rake receiver as well as to a chip-level equalizer in a Rayleigh fading channel in Section IV, followed by concluding remarks in Section V.

II. SYSTEM MODEL

Since the downlink is considered, synchronous transmission of all signals through the same multipath channel is assumed. The complex envelope of the received signal at

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user terminal can be written as

$$r(t)=\sum_{k=1}^{K}\sum_{m=0}^{M-1}A_k b_k^{(m)}\sum_{l=1}^{L}c_l(t)s_k^{(m)}(t-mT-\tau_l)+n(t),$$

where K is the number of users, M is the length of observation window in symbols, L is the number of paths, A_k is the average received amplitude of kth user, $b_k^{(m)}$ is the mth symbol of kth user, $c_l(t)$ is the time-variant complex channel coefficient of lth path, $s_k^{(m)}(t)$ is the spreading waveform of mth symbol of kth user given by convolution of spreading sequence and chip waveform, T is symbol interval, τ_l is the delay of lth path and $n(t)$ is complex white Gaussian noise process.
The discrete-time received signal can be written as

$$\boldsymbol{r}=\sum_{k=1}^{K}\boldsymbol{D}\boldsymbol{C}\boldsymbol{A}_k\boldsymbol{S}_k\boldsymbol{b}_k+\boldsymbol{n}\in\mathbb{C}^{MN_cN_s}, \tag{1}$$

where N_s is the number of samples per chip, N_c is the spreading factor, $\boldsymbol{D}=[\boldsymbol{d}_1^{(1,0)},\ldots,\boldsymbol{d}_L^{(1,0)},\boldsymbol{d}_1^{(2,0)},\ldots,\boldsymbol{d}_L^{(N_c,0)}\boldsymbol{d}_1^{(1,1)},\ldots,\boldsymbol{d}_L^{(N_c,M-1)}]\in\mathbb{R}^{MN_CN_s\times LMN_c}$ is a path delay and chip waveform matrix with column vector $\boldsymbol{d}_l^{(n,m)}\in\mathbb{R}^{MN_cN_s}$ containing samples from appropriately delayed chip waveform[1]. $\boldsymbol{C}=\operatorname{diag}\left(\boldsymbol{c}^{(1,0)},\ldots,\boldsymbol{c}^{(N_c,0)},\boldsymbol{c}^{(1,1)},\ldots,\boldsymbol{c}^{(N_c,M-1)}\right)\in\mathbb{C}^{LMN_c\times MN_c}$ is a block diagonal channel matrix with column vector $\boldsymbol{c}^{(n,m)}\in\mathbb{C}^{L}$ containing channel coefficients for L paths. $\boldsymbol{A}_k=A_k\boldsymbol{I}_{MN_c}$ contains the average received amplitudes and $\boldsymbol{S}_k=\operatorname{diag}(\boldsymbol{s}_k^{(0)},\cdots,\boldsymbol{s}_k^{(M-1)})\in\mathbb{C}^{MN_c\times M}$ is a block diagonal spreading sequence matrix with column vector $\boldsymbol{s}_k^{(m)}$containing the spreading sequence for the kth user's mth symbol. The cell specific scrambling sequence is included to the spreading sequence, and the sequences are normalized so that $\boldsymbol{S}_k^{\mathrm{H}}\boldsymbol{S}_k=\boldsymbol{I}$. Vector $\boldsymbol{b}_k=[b_k^{(1)},\cdots,b_k^{(M)}]^{\mathrm{T}}$ contains the transmitted symbols of kth user, and $\boldsymbol{n}\in\mathbb{C}^{MN_cN_s}$ contains samples from white complex Gaussian noise process with covariance $\sigma_n^2\boldsymbol{I}_{MN_cN_s}$.

III. RECEIVERS

In this section, the conventional Rake receiver is shortly defined, followed by the definition of the chip-level LMMSE equalizer in Section III-A. Necessary approximations made for the LMMSE equalizer are presented, and the symbol-level equalizer is proposed in Section III-B. Finally an adaptation method for the symbol-level equalizer is presented in Section III-C.
With the introduced system model, the decision variable of the Rake receiver for arbitrary user 1 is given by $\boldsymbol{y}_{[R]}=\boldsymbol{S}_1^{\mathrm{H}}\left(\boldsymbol{D}\boldsymbol{C}\right)^{\mathrm{H}}\boldsymbol{r}$, i.e., the received signal is filtered by the chip waveform, appropriately time-aligned and weighted with channel coefficients in the Rake fingers, coherently combined and finally despread. The abnormal order of the maximal ratio combining (MRC) and despreading has no effect on the receiver performance due to the linearity of operations.

A. LMMSE equalizer

The chip-level LMMSE equalizer is obtained by minimizing the mean square error between the equalizer output and the total transmitted signal, i.e., by solving

$$\boldsymbol{w}=\arg\min_{\boldsymbol{w}}\mathrm{E}\left[|\boldsymbol{w}^{\mathrm{H}}\boldsymbol{r}-\sum_{k=1}^{K}\boldsymbol{A}_k\boldsymbol{S}_k\boldsymbol{b}_k|^2\right], \tag{2}$$

where minimization is carried out elementwise. The optimization problem in (2) can be solved, resulting [11]

$$\boldsymbol{w}=\left(\boldsymbol{D}\boldsymbol{C}\left(\sum_{k=1}^{K}A_k^2\boldsymbol{S}_k\boldsymbol{S}_k^{\mathrm{H}}\right)\boldsymbol{C}^{\mathrm{H}}\boldsymbol{D}^{\mathrm{H}}+\sigma_n^2\boldsymbol{I}\right)^{-1}\boldsymbol{D}\boldsymbol{C}. \tag{3}$$

The standard symbol-level LMMSE equalizer [3]-[4], feasible for DS-CDMA systems employing short spreading sequences, is defined for the user 1 by

$$\boldsymbol{L}=\arg\min_{\boldsymbol{L}}\mathrm{E}\left[|\boldsymbol{L}^{\mathrm{H}}\boldsymbol{r}-\boldsymbol{b}_1|^2\right]. \tag{4}$$

It can be shown that the decision variables of chip-level and symbol-level LMMSE receivers are equal up to a scalar for a given observation window [10]. This is an expected result, since LMMSE estimator commutes over linear (or affine) transformations [11], like despreading. The equal performance of LMMSE receivers was verified numerically in [12].
As seen from (3), the LMMSE solution of chip-level equalizer depends on the spreading sequences of all users with the period of long scrambling code. This follows from the dependency between consecutive chips to be estimated. The optimal solution changes from chip to chip, and an adaptive chip-level equalizer will not reach the exact optimal tap coefficients. The adaptive versions of chip-level LMMSE equalizer are build on simplifications that (*i*) spreading sequences are random, and that (*ii*) the random spreading sequences are white and independent from the spreading sequences of other users. The simplifications are unavoidable due to the non-stationarity of the LMMSE solution in (3). The LMMSE chip-level equalizer derived according to the simplifying approximations is given by [10]

$$\tilde{\boldsymbol{w}}=\left(s^2\sum_{k=1}^{K}A_k^2\boldsymbol{D}\boldsymbol{C}\boldsymbol{C}^{\mathrm{H}}\boldsymbol{D}^{\mathrm{H}}+\sigma_n^2\boldsymbol{I}\right)^{-1}\boldsymbol{D}\boldsymbol{C}, \tag{5}$$

where s^2 denotes the square value of chip. The decision variable after the correlation with spreading sequence is given by

$$\begin{aligned}\boldsymbol{y}_{[\tilde{w}]}&=\boldsymbol{S}_1^{\mathrm{H}}\tilde{\boldsymbol{w}}^{\mathrm{H}}\boldsymbol{r} \qquad (6)\\&=\boldsymbol{S}_1^{\mathrm{H}}\boldsymbol{C}^{\mathrm{H}}\boldsymbol{D}^{\mathrm{H}}\left(s^2\sum_{k=1}^{K}A_k^2\boldsymbol{D}\boldsymbol{C}\boldsymbol{C}^{\mathrm{H}}\boldsymbol{D}^{\mathrm{H}}+\sigma_n^2\boldsymbol{I}\right)^{-1}\boldsymbol{r}.\end{aligned}$$

[1]In the notation $l=1,2,\ldots,L$ refers to the path, $n=1,2,\ldots,N_c$ to the chip, and $m=1,2,\ldots,M$ to the symbol.

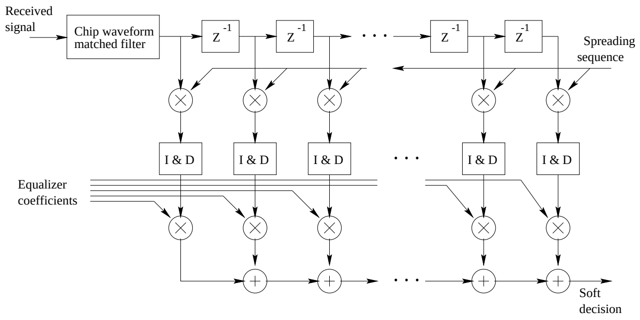

Fig. 1. Structure of symbol-level equalizer.

The performance degradation caused by the aforementioned simplifications can be assessed by comparing the performances of the equalizers given in (3) and (5). The degradation caused by the approximation is relatively minor [12]-[14], and the equalizer in (5) is well-suited for adaptive implementations.

B. Symbol-level channel equalization

The main benefit of the equalizer in (5) is the independency from the scrambling code while maintaining the most of the performance gains provided by the LMMSE equalization. The independency from the scrambling code allows the use of simple adaption algorithms with the equalizer. A serious drawback of the equalizer is its high complexity due to the equalization at chip level.

A decision variable in (6) can be written as

$$\begin{aligned} y &= \sum_{g=0}^{N_c-1} s^*(n+g) \sum_{l=-L}^{L} \tilde{w}_l^* r(n+g+l) \\ &= \sum_{l=-L}^{L} \tilde{w}_l^* \sum_{g=0}^{N_c-1} s^*(n+g) r(n+g+l), \end{aligned} \tag{7}$$

where $(\cdot)^*$ denotes complex conjugate, $s(n)$ spreading code, $\tilde{w}_l$ the coefficients of the equalizer with $2L+1$ taps, and $r(n)$ denotes the received signal samples after chip waveform filtering.

From (7) it is easily noted that the equalization can be performed at symbol-level by replacing the single correlator following the chip-level equalizer with a bank of correlators. The receiver structure is depicted in Fig. 1. Due to increased number of correlators, no complexity reduction is obtained with respect to the number of additions. However, the number of multiplications due to the equalization is reduced by the order of spreading factor N_c.

C. Adaptive symbol-level equalizer

The equaliser coefficients $\tilde{\boldsymbol{w}}$ can be calculated by solving

$$\tilde{\boldsymbol{R}} \boldsymbol{w} = \tilde{\boldsymbol{p}}, \tag{8}$$

where $\tilde{\boldsymbol{R}}$ is the estimate of the received signal's covariance matrix, and $\tilde{\boldsymbol{p}}$ is the channel response estimate. The direct calculation of the coefficients may appear unnecessarily complex at the first sight. However, the covariance matrix has a Toeplitz structure due to chip rate sampling, and thus (8) can be efficiently solved with Levinson algorithm [15]. Also the equaliser coefficients need to be updated only few times in a slot. These two aspects reduce the computational burden of (8) to a reasonable level. The use of Levinson algorithm was independently proposed for the chip-level equalizers in [9] and [10].

The channel response can be estimated with the common pilot channel (CPICH). The covariance matrix $\boldsymbol{R}$ is completely defined by the zero and positive lag values of the received signal's autocorrelation function. The most straightforward way to estimate received signal's autocorrelation is a moving average

$$\tilde{\boldsymbol{a}}_0 = 1/J \sum_{j=1}^{J} r(n-Mj) \otimes \begin{pmatrix} r(n-L-Mj)^* \\ r(n-L+1-Mj)^* \\ \vdots \\ r(n+L-Mj)^* \end{pmatrix},$$

where $\otimes$ is Kronecker product. Parameter M controls the sample rate on autocorrelation estimation, e.g., with $M=8$, the multiplications are done every eight sample. The conjugate symmetry properties of autocorrelation function can be utilized to improve the estimate by

$$\tilde{\boldsymbol{a}} = (\tilde{\boldsymbol{a}}_0 + \boldsymbol{E}\tilde{\boldsymbol{a}}_0^*)/2, \tag{9}$$

where $\boldsymbol{E}$ is the counter-identity matrix.

IV. NUMERICAL RESULTS

To obtain a good understanding of the potential performance that the presented receiver can offer, bit error rates (BER's) were evaluated in a Rayleigh fading frequency-selective channel. QPSK modulation was used employing root raised cosine pulses with roll-off factor of 0.22. Random cell specific scrambling code and Walsh channelization codes were used. Spreading factor 64 was used on the common pilot channel, and the transmission power of pilot channel was scaled to be 10% from the total transmitted power. The cell contained 4 users with spreading factor 8 and a desired user with spreading factor 64. The powers of users were scaled to produce the same E_b/N_0 value to all users. The chip rate was set to 3.84 Mchip/s corresponding to 260 ns chip interval. BER's were evaluated in a 4-path channel with average relative path powers of 0, -1, -9 and -10 dB, and path delays of 0, 309, 716 and 1090 ns (resembling ITU's vehicular channel model A). The terminal speed was assumed to be 60 km/h.

The received signal was modelled with two samples per chip, and the output of chip waveform matched filter was sampled at chip rate. In the simulations the Rake receiver contained fingers allocated at correct path delays. Equalizers had 23 taps. Channel response was estimated with CPICH and a moving average filtering.

In Fig. 2, BER's are presented for the conventional Rake receiver, chip-level equalizer and symbol-level equalizer, both using Levinson algorithm and estimated channel

TABLE I

Computational requirements of the equalizers.

	# of additions	# of multiplications
Chip-level equalizer	413 Mops	406 Mops
Symbol-level equalizer	408 Mops	58 Mops

response and autocorrelation in the adaptation of equalizer coefficients. The chip-level equalizer used every received signal sample in the autocorrelation estimation ($M = 1$), and the symbol-level equalizer used every 8th sample ($M = 8$). The equalizer coefficients were recalculated either at rate 15 kHz or at rate 3.75 kHz. A significant performance gain provided by the equalizers over the Rake receiver can be noted from the figure, as well as almost negligible performance difference between the chip-level and symbol-level equalizers.

For the aforementioned case, the receivers' requirements for real-number additions and multiplications are tabulated in Table I. The table contains only the computational load of correlations with the spreading sequence, equalization, autocorrelation estimation and Levinson algorithm. E.g. the effect of channel estimation, or chip waveform filtering is excluded. Additionally, the equalizers involve one division per equalizer tap at every equalizer update. For the values presented on the table, the equalizer coefficients of both chip-level and symbol-level equalizers were updated at rate 3.75 kHz, and $M = 8$ was employed. The algorithmically simple autocorrelation estimation employed in the paper forms over 70% of the multiplications in the symbol-level equalizer.

V. CONCLUSIONS

One approach to improve the performance of WCDMA downlink receivers was studied in this paper, namely channel equalization at the symbol level. The introduced symbol-level equalizer is based on same approach as the chip-level equalizer, i.e., to restore to some extent the orthogonality of users, and, thus, suppress MAI when orthogonal spreading sequences are employed. However, by performing the equalization with symbol level signal processing instead of chip level processing, significant reduction in the required number of multiplications is achieved without compromising the performance. The performance of the equalizers was numerically evaluated and compared to the performance of the Rake receiver in a Rayleigh fading multipath channel. The results show significant performance improvements when equalizers are employed instead of the conventional Rake receiver, and that both symbol-level and chip-lever equalizers obtain equal performance.

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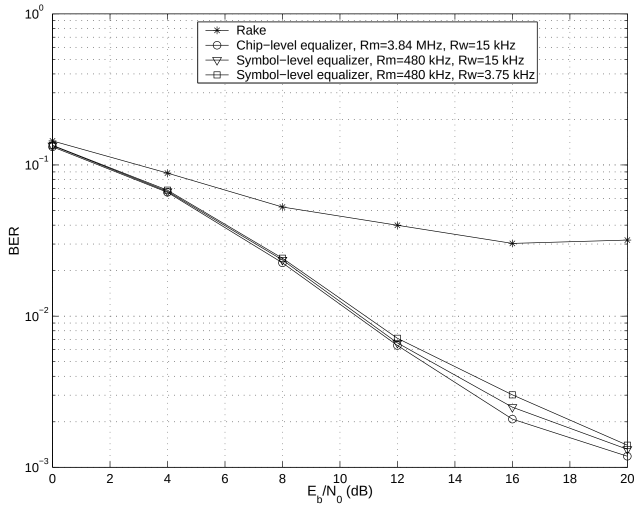

Fig. 2. BER vs. E_b/N_0 for Rake receiver, chip-level and symbol-level equalizers with different equalizer coefficient update rates.

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